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An Automatically Calibrated Universal Measuring Equipment for Noise in Semi-Conductors.

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ABSTRACT

Recent developments in X and Gamma Ray spectrometry and Infra-red detection require the use of Junction Field-Effect Transistors (JFets) which exhibit very low levels of noise in the frequency range from a few Hertz to 100KHz. ~~In many applications the optimum~~ performance is achieved ~~by cooling the detector in a cryostat, and the~~ detector and JFet may operate at a temperature as low as 80°K.

~~It has been~~ shown that the low frequency noise of JFets is strongly ~~dependent on device operating parameters~~ and can vary greatly between ~~similar devices operating~~ under the same conditions. In order to investigate the sources of low frequency noise further, and to select JFets for use with detector systems, it has been necessary to accurately measure the low frequency noise performance of devices over a wide range of ~~operating conditions~~.

A micro-computer controlled equipment has been developed which can measure the noise of Fets, NPN and PNP bipolar transistors in the frequency range 10Hz to 100KHz, to a greater accuracy than that obtainable from available equipment. The device operating parameters may be varied under computer control, allowing noise measurements to be performed automatically over a range of conditions. When measuring four-terminal (or tetrode) Fets, the substrate voltage may be varied independently of the gate voltage.

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ILLUSTRATIONS

1. INTRODUCTION

High resolution X and Gamma ray Spectrometry and Infra-red detection systems require a pre-amplifier with a low noise transistor input to amplify the detector signal before it is processed by a pulse-shaping network to optimise the signal-to-noise ratio (SNR). The noise of the input transistor usually limits the SNR that may be achieved, and the nature of the signal processing method employed is such that capacitance and leakage current at the input to the pre-amplifier both degrade the SNR. In order to achieve low noise with low capacitance and low leakage small gate area JFets are used. In some applications the optimum SNR is obtained by processing the detected signal for time periods up to 10ms or more, and under these conditions it is the noise of the input JFet at frequencies down to a few Hertz which limits the SNR.

It has been shown (1,2,3) that the most significant source of low frequency noise in small gate area JFets is the drain current modulation caused by charge transitions in single point defects situated close to the edge of the channel and near the pinch-down region of the device. The magnitude of the low frequency noise is critically dependent on device operating conditions and may show large peaks as a function of temperature, drain current and substrate voltage. An example of the variation of low frequency noise with substrate voltage and temperature is shown in Figure 1. This illustration demonstrates the need to undertake many noise measurements over a range of conditions in order to establish the operating point for optimum noise performance.

A fully automatic noise measuring equipment has been developed for measuring the noise of Fets in the frequency range 1GHz to 100KHz and at temperatures between 79^{o}K and 350^{o}K. Drain and substrate voltages may be varied between 0V and 10V and drain current between 100μA and 100mA. The equipment may also be used for measuring the noise of NPN and PNP bipolar transistors. The device operating conditions are accurately defined and devices may be measured with noise levels between 0.3 nVolts per root Hertz and 2μVolts per root Hertz, (see Note 1). The transistor operating parameters and execution of the noise measurements are controlled with a micro-computer. The computer provides a digital read-out of the noise measurement, eliminating the common problem of human error associated with reading a meter. Some typical noise measurements performed with the equipment are given in Appendix A.

2. GENERAL DESCRIPTION

The general problem in the measurement of noise in amplifying devices is that of knowing the magnitude of the gain and bandwidth of the following amplifiers and filters. The noise of devices is required to be known at particular frequencies. However, filters of finite bandwidth must be used and the difficulty arises in determining their actual effective bandwidth. These problems are overcome here by the use of a calibrated and very stable noise source. Following the measurement of the amplitude of the noise of the device under test, the equipment is calibrated by performing a second measurement after introducing a well defined white noise source whose amplitude is very much greater than the noise of the device. The actual magnitude of the device noise is then obtained by solving a simple equation.

NOTE 1: It is common practice to measure noise power per unit bandwidth, and therefore noise voltage is measured per root frequency.

In this equipment the device under test forms the input stage of an amplifier. The remainder of this first amplifier is designed to have a very low and well defined noise performance, allowing noise sources in the amplifier to be subsequently accounted for when the noise of the device is finally calculated in the micro-computer. To prevent extraneous noise from corrupting the measurement, the first and second stage amplifiers are contained in a fully screened free standing module (figure 2).

The device noise after considerable amplification is passed to various filters of different frequencies between 10Hz and 100KHz. The filtered noise signal is rectified and smoothed and is used to generate a pulse-train with a frequency proportional to the rms magnitude of the noise signal. The pulses are counted in scalers for a defined time period.

A block diagram of the noise measuring equipment is shown in figure 3.

3. PRE-AMPLIFIER

The pre-amplifier controls the operating conditions of the device under test and amplifies the noise signal typically over the 3dB frequency band 1Hz to 1MHz.

In order to achieve the best possible noise performance whilst maintaining a rapid response to changes in device operating conditions, it has been necessary to provide separate pre-amplifiers for measuring the noise of N-channel Fets, P-channel Fets, NPN and PNP transistors. Noise measurement accuracy is maintained over a range of drain or collector currents by optimising the design of each pre-amplifier for a particular current range, and separate units are provided for operating over different current ranges. A discussion of the limitations to the drain current range is given in Appendix B.

The device operating conditions are determined by well defined and stable voltage levels (parameter control signals) which are generated by the Transistor Parameter Controller and transferred to the pre-amplifier by the Pre-amplifier Controller, along with other control signals and power supplies.

The design of each pre-amplifier is broadly similar; for example, the N-channel and P-channel units are nearly identical except for a voltage polarity change in the first stage amplifier and the parameter control circuits. The bipolar pre-amplifiers differ from the Fet units only in the method for controlling the base voltage of the device under test. Each pre-amplifier has a unique identity code which can be read by the micro-computer. The N-channel pre-amplifier only is described here, and the base voltage control in the NPN pre-amplifier is discussed briefly in a later section.

3.1 N-Channel Pre-amplifier

A block diagram of the N-channel pre-amplifier is shown in figure 4. The device under test, T1, is at the input of a preliminary amplification circuit incorporating a cascode transistor T2, source-follower T3 and feed-back components C1, R1 and R2. The substrate voltage, drain voltage and drain current are precisely controlled and the gate voltage is automatically biased to the correct operating voltage. The first stage gain is determined by R1 and R2, and is dependent on the drain current range of the pre-amplifier. The open loop gain of the first stage amplifier is proportional to the mutual conductance of the device which increases approximately as the square root of the drain current. To ensure

that the demanded gain is always less than the open loop gain, the demanded first stage gain has been reduced in pre-amplifiers which operate over a low current range. The reduction of signal gain at low currents is advantageous as the noise of devices generally increases as the drain current decreases.

The cascode current, Ic, is generated from a low noise constant current circuit. To maintain the accuracy of the drain current setting at low drain currents, Ic is chosen to be smaller than the lower limit of the current range of the pre-amplifier, but sufficiently large for T2 to have a good frequency response and low emitter resistance noise.

The first stage of amplification is followed by a second stage wide-band discrete amplifier, A1, with a low noise Fet input. The two stages provide a total gain of between 2,500 and 15,000 depending on the current range. A voltage comparator at the output of A1 signals when the amplifier circuit is temporarily out of balance, which may occur immediately after changing the operating parameters of T1, for example.

3.1.1 Parameter Control Circuits

An important feature of the noise measuring equipment is its ability to rapidly change the device operating conditions between noise measurements, while maintaining a high level of noise filtering in the parameter control circuits during the measurement period. This has been achieved by incorporating a dynamic filter network using an Operational Transconductance Amplifier (OTA) in each of the parameter control circuits. The OTA is used with a feedback loop which supplies current to a capacitor, adjusting the voltage on the capacitor until the circuit is balanced. The OTA and capacitor provide a filter network with a time constant dependent on the gain Ao of the OTA, which can be varied by applying an external bias current.

During the noise measurement period Ao is sufficiently small for the noise at the input to the OTA to be adequately filtered. Following a change in device operating conditions Ao is increased, shortening the filter time constant and allowing the parameter control circuit to respond rapidly.

A schematic diagram of the drain volage control circuit is shown in figure 5. The drain voltage control signal is filtered by A1 and C1 and generates a well defined voltage at point P1. Transistor T2 provides a voltage drop equal to the base-emitter voltage of the cascode transistor, setting the drain voltage of the device under test to be equal to the voltage at P1. The gain of A1 is varied by applying a control signal, (the Initialise signal), to change the current into the bias input of A1.

A similar circuit is provided for setting the substrate voltage, but to minimise the injection of noise into the Fet, the capacitor and substrate are connected directly to the output of the OTA, which is wired as a simple voltage follower. The gain of the OTA determines the maximum current which it may supply, thus setting a maximum limit to the substrate leakage current which may be allowed.

Figure 6 is a schematic diagram of the gate voltage control circuit. OTA A1 adjusts the gate bias of the device under test T1 until the inverting input of the second stage amplifier is at zero volts. The gain of A1 during the noise measurement period determines the maximum allowed gate leakage current. An additional OTA, A3, and buffer amplifier A2 provide a protect circuit which can clamp the gate of T1 to zero volts.

During normal operation the gain of A3 is zero and the gate of T1 is set at its operating voltage. When the device under test is changed, a control signal (the Protect signal) is automatically triggered and increases the gain of A3, driving current into C1 until the gate voltage has reached zero. A similar protection circuit is incorporated in the substrate control circuit, and the device may be removed or inserted with power on without causing damage.

A schematic diagram of the drain current control circuit is shown in figure 7. To allow the drain current to be varied over as large a range as possible while maintaining noise measurement accuracy, it has been necessary to generate the drain current supply from a +60V supply line. A detailed discussion of noise sources and limitations to the drain current dynamic range is given in Appendix B.

The problem of coupling the OTA amplifier, A1, to the +60V line has been overcome by using opto-couplers, OT1 and OT2. The opto-couplers supply current to capacitor C1 which adjusts the voltage at point P2 until sufficient current flows through R4 to balance the circuit. The drain current is supplied through transistors T4 and T5 and is determined by the magnitude of R6 and R7. Resistor R8 supplies a nearly constant current Ic which is approximately equal to the cascode current. The magnitude of Ic is dependent on the drain voltage to a small degree, and in order to reduce the error in the drain current setting to less than 1% the micro-computer applies a small correction to the drain current which takes into account variations of Ic with drain voltage.

3.1.2. Noise Considerations

The essential requirement of a circuit for amplifying noise is that the contribution to the output noise signal from elements in the circuit is small compared to the contribution from the device under test, and that these additional noise souces are well defined and stable and may be subtracted from the total noise measurement.

The most significant excess noise contributions are from noise sources which are introduced directly into the device under test or into the first stage of amplification. These are, with reference to figure 4,

(i) the thermal noise of resistor R1,

(ii) the emitter resistance and base spreading resistance of the cascode transistor, T2.

(iii) the base current noise of T2.

(iv) noise from the parameter control circuits.

R1 is a highly stable metal film resistor with a accurately known value of between 5Ω and 10Ω, depending on the current range of the pre-amplifier. The value of R1 is reduced in the higher current pre-amplifiers, and the first stage gain subsequently decreased, in order to match the lower noise and higher mutual conductance of the device under test. A subtraction of the noise of R1 is performed in the noise measurement calculation.

The requirements of the cascode transitor are that it must possess a high forward current gain (Hfe) and low base and emitter resistance noise at the demanded cascode current. The characteristics of the cascode transistor are matched to the drain current range of the pre-amplifier, and

devices may be found which contribute less than 1% to the noise measurement of the device under test.

The substrate voltage, gate voltage and drain voltage control circuits provide sufficient noise filtering for their contribution to the total noise signal to be negligible. A discussion of noise sources in the drain current supply circuit and limitations to the drain current dynamic range is given in Appendix B. In brief, the magnitude of the excess noise injected into the drain of the device under test is dependent on the noise in the fixed resistor through which the current is supplied, and the noise in the voltage supply line.

In order to limit the error in the noise measurement to a few percent while providing a drain current dynamic range of at least 10:1 in each pre-amplifier, it has been necessary to generate the drain current from a +60V supply line. A sufficiently low noise voltage supply has been obtained by designing a mains operated unit which incorporates choke-input filters and sophisticated stabilisation circuits. The power supply is discussed in a later section. Additional filtering of the voltage supply is provided in the pre-amplifier, and the noise contribution from the +60V supply is negligible. However, a small correction of a few percent is made in the noise measurement calculation to subtract the noise of the fixed resistor which supplies the drain current.

3.1.3 Calibration

The noise measuring equipment is calibrated by introducing a very stable and precisely defined white noise source into the input of the pre-amplifier after each noise measurement. This technique of automatic calibration eliminates the effect of small variations in the gain of the pre-amplifier which may occur as a result of changes in device operating conditions, as well as removing the need to establish the exact gain and bandwidth of the filter circuits.

A white noise source is generated using a pseudo-random sequence generator whose amplitude is accurately known in the range 10Hz to 100KHz. Generation of the calibrated noise signal is discussed in a later section.

Two modes of calibration are provided. With reference to figure 4, the first method involves introducing the calibration signal through a stable resistor network, R1 and R2, into the gate of the Fet under test. R1 and R2 attenuate the calibration signal by an accurately known factor, and the noise measurement is calibrated in terms of the equivalent noise voltage, En, referred to the gate of the device.

In some applications it is more convenient to refer to the equivalent noise current Ino in the drain, as the thermal noise current of Fets is almost independent of substrate voltage and temperature. Methods for measuring Ino usually rely on measuring En and the mutual conductance of the device separately. Here, the noise current measurement may be directly calibrated by introducing the calibration signal through an accurately defined and stable resistor, R5, into the drain of the device.

The required amplitude of the calibration signal is automatically calculated and controlled by the micro-computer. Calculation of the required amplitude for En measurements is straight forward, since the gain from the input of the pre-amplifier is almost independent of device operating conditions and is known approximately. Calculation of the calibration signal amplitude for Ino measurements requires a knowledge of the mutual conductance, gm, of the device. Gm varies as the operating

conditions are changed, and it is necessary to perform a trial noise measurement (using a very short integration time) prior to each Ino measurement, in order to estimate the calibration signal amplitude required.

3.2 Base Voltage Control in the Bipolar Transistor Pre-amplifier.

The only significant difference in the circuit design of the Fet pre-amplifiers and the bipolar transistor pre-amplifiers is the method for controlling the base voltage of the transistor under test. The base voltage requires to be set only over a limited range between 0.5V and 0.8V, and the controlling circuit must be capable of supplying a current of up to a few milli-amperes into the base of the transistor. The magnitude of any noise which is introduced into the base must be small and known accurately, so that a correction to the noise measurement can be made.

The base voltage is controlled by applying a variable current, Io, into an accurately defined and stable fixed resistor of a few Ohms. Io has two components. The first component is a fixed, very low noise current which is supplied from the 60V line after extensive filtering. Secondly, a variable current is supplied whose magnitude is controlled by an OTA which adjusts the magnitude of Io to set the correct operating base voltage. The gain of the OTA may be varied to provide a dynamic filter network in a similar way to the gate voltage control circuit.

The base voltage may be varied between 0.5V and 0.8V with only the noise of the fixed resistor being added to the noise measurement of the device under test.

4. FILTER INTERFACE

In order to maximise the noise signal dynamic range in the equipment, it is desirable to adjust the signal amplitude at the output of the pre-amplifier to match the dynamic range of the filter units. This has been achieved by designing a Filter Interface which provides variable signal gain, controlled by the micro-computer, between the pre-amplifier and the filter units.

A transient voltage swing at the input to the filter units may cause the low frequency filter circuits to saturate, requiring many seconds for the noise signal to return to its correct level. Such voltage swings may occur at the output of the pre-amplifier after a change in device operating conditions. To allow noise measurements to begin within a few seconds of changing the device operating conditions, the Filter Interface has been provided with a facility for isolating the input of the filter units from the output of the pre-amplifier.

A schematic diagram of the filter interface is shown in Figure 8.

4.1 Isolation of the Noise Signal from the Filters

Isolation of the noise signal from the input to the filter units is achieved by the use of an OTA, A2 in figure 8. The gain of A2 is determined by the bias current, Ib, supplied through R12 and D1, and may be changed by a logic control signal. During normal operation the gain of A2 is set at a high value and the source of transistor T1 (point P2) follows the mean signal input level at P1.

The signal input is taken to buffer amplifier A1 whose output (point P1) can be connected by an Fet analogue switch AS1 to one of two positions. During normal operation the signal at P1 is AC-coupled (C1,R1) into the input of amplifier A3. When isolation is required, the gain of A2 is reduced to zero and analogue switch AS1 connects P2 to C1. P2 maintains the original mean signal level at P1 to within 1mV for a period of many tens of seconds, decaying with a very long time constant determined by the value of C3 and the gate leakage current of T1. When the output of the pre-amplifier has settled to its correct operating level, the signal path may be re-connected by AS1 and the gain of A2 increased.

4.2 Computer Controlled Variable Gain

Variable gain is provided by attenuating the noise signal using a resistor network, R2 to R5. Each of the resistors R3, R4 and R5 can be shorted to the ground 'y a bipolar transistor analogue switch, each switch being set by a control code determined by the micro-computer. The resistor network provides eight possible gain settings over a range of approximately 10:1. The gain at each setting need not be defined precisely as it remains fixed for both the noise and calibration signals. The micro-computer calculates the required gain setting by performing an initial trial noise measurement before measuring the device noise accurately.

In order to provide an accurate calibration of the signal gain in the pre-amplifier and filter units, it is important that the magnitude of the calibration signal is very much greater than the noise from the device under test. To allow the use of a large calibration signal while keeping the signal within the dynamic range of the filter units, the gain in the Filter Interface is reduced by an accurately determined factor before the calibration signal is applied. The gain determined by amplifier A4 may be reduced by operating an analogue switch AS5. R7, R8 and R9 are accurately defined and stable resistors. The collector-emitter capacitance of the grounding transistor in AS5 is sufficiently large for the collector impedance to ground to be significant at the highest frequency of interest (100KHz). Therefore, to achieve a gain change defined to better than 1%, resistor R7 is connected in parallel with AS5.

The physical nature of noise in most semi-conductor devices is such that the magnitude of the noise signal, measured in the range 10Hz to 100KHz, decreases as the frequency increases. The noise signal amplitude in filter units with centre frequency, Fc, greater than 10KHz might therefore be expected to be less than in lower frequency filters. However, the operating dynamic range of the high frequency filters is limited at low signal levels, and it is therefore desirable to have a greater signal gain to the input of filter units with Fc greater than 10KHz. For this reason a second output from the Filter Interface is provided following further amplification by A5 and A6. The gain of A5 and A6 is accurately defined and is switched to the low gain setting during the calibration measurement, using analogue switches AS6 and AS7. Resistors R10 and R11 provide a small attenuation to give unity gain between the two outputs when A5 and A6 are in the low gain setting.

5. FILTER UNITS

After amplification by the Filter Interface the noise signal is processed by a number of parallel filter channels, each one of which incorporates a bandpass filter of specified centre frequency, Fc, and 3dB bandwidth of Fc/N, where N is between 2.5 and 10. The ratio of bandwidth to centre frequency is increased at low frequencies to increase the

statistical accuracy of the noise measurement. The noise component of the signal within the bandwidth of the filter is rectified, smoothed and converted into a pulse train of frequency proportional to the rms signal amplitude.

Figure 1 is a schematic diagram of a filter channel. Amplifiers A1 and A2 provide initial amplification before the noise signal is filtered by bandpass filter F. The signal path is AC-coupled up to the input of A4 to eliminate the off-set voltage drift in the amplifiers. A1 and F are preceded by low-pass filters (R1, C1 and R2,C2) with a corner frequency of approximately 8Fc. The low-pass filters remove the high frequency components of the noise signal which may otherwise saturate the amplifiers or the band-pass filter.

The range of noise measurement which may be performed has been increased by varying the gain of amplifier A2 in the low frequency filter units under computer control. A feedback resistor on A2 may be selected using an analogue switch, allowing the gain to be reduced to deal with very large noise signals which may occur at low frequencies.

The signal gain following F is fixed for all filter channels. Amplifiers A4 and A5 and diodes D1 and D2 form a linear rectification circuit. The two halves of the rectified signal at P2 and P3 are smoothed and summed by A7, with a smoothing time-constant of at least 20/Fc, before being taken to the Voltage-to-Frequency Converter (VFC). The gain of A7 is approximately five, which is greater than the Form Factor of the unsmoothed signal at P2 and P3. Thus for a random signal the full-scale output of A7 and the VFC may be achieved while the noise peaks are still in the linear range of A5 and A6.

The VFC generates a pulse train with frequency proportional to the signal amplitude. The noise signal is averaged by counting the pulses in scalers contained in the Micro-computer Interface for a specified integration time. The Micro-computer Interface is a modified 6161 MOUSE module.

5.1 Operating Dynamic Range

The linearity of the signal response, and therefore operating dynamic range of the filter units, is limited by the rectification stage. The DC off-set drift of A4, A5 and A6 produces a non-linear response at small signal levels. In addition, in high frequency filters the finite slew rates of A5 and A6 can introduce distortion into the rectified, but unsmoothed, signals at P2 and P3. Non-linearity of less than 1% can be achieved over a dynamic range of 100:1 for Fc less than 10KHz, falling to 10:1 for Fc of 100KHz. However, the extra switched gain facility for filters with Fc greater than 10KHz provided in the Filter Interface allows the effective dynamic range of the 100KHz filter to be 100:1.

The additional variable reduction in gain in filters of frequency less than 1KHz allows measurements to be performed on devices whose noise may vary as a function of frequency, at any one operating condition, by a factor of 100 at frequencies between 1KHz and 100KHz and by a factor of 1000 at frequencies less than 1KHz.

5.2 Statistical Accuracy of the noise measurement.

It is well known that a random distribution of N numbers about some mean value has a standard deviation from the mean which is proportional to $1/(N)^{\frac{1}{2}}$. Similarly, the statistical accuracy of the noise measurement

varies with the total number of pulses, No, generated by the VFC's and counted in the scalers. No is dependent on the product of the integration time and the filter bandwidth, and it has been found that the standard deviation of the amplitude measurement of a random noise input signal can be approximately determined using the following empirical formula,

$$Sd\ (\%) = 50/(T.Fc)^{\frac{1}{2}}$$

where Sd = Standard Deviation (%)
T = Integration Time (Secs)
Fc = 3dB Bandwidth of Filter (Hz)

To obtain adequate statistical accuracy in an integration period of only a few seconds, the 3dB bandwidths have been chosen to be 10% of the filter centre frequencies of 1KHz or greater, increasing to 40% for the 10Hz filter. For example, a 10 second integration period will achieve a standard deviation in the measured noise amplitude of approximately 1.6% at 1KHz and 8% at 10Hz.

6. CALIBRATED NOISE SOURCE

The requirements of the calibrated noise source are to provide an accurately defined and stable white noise source for use in the frequency range 10Hz to 100KHz, the magnitude of which is defined to better than 1% over a dynamic range of 100:1.

It is well known (4,5) that a pseudo-random sequence generator may be formed using shift registers with feedback via an exclusive OR gate from two specific outputs of the registers. Figure 10 is a schematic diagram of such a circuit. The shift register is clocked from an accurate and stable crystal generator. In order to provide a calibrated noise source which exhibits a sufficient degree of randomness for this application, it has been necessary to mix the outputs of three pseudo-random generators containing shift registers of specific lengths. Such a circuit has been reported by White(6).

The output of the combined pseudo-random sequence generators is passed through a level changer whose amplitude is accurately controlled by the micro-computer using a Digital-to-Analogue Converter. The amplitude is defined to better than 1% over a dynamic range of 100:1.

It has been shown by White(6) that the magnitude Eo of the calibrated noise source at frequency F is given by the following formula;

$$Eo = (A/2Fo)^{\frac{1}{2}}\ \mathrm{Sin}\ (\pi F/Fo)\ /\ (\pi F/Fo)$$

$$\approx (A/2Fo)^{\frac{1}{2}} \text{ for } Fo \gg F$$

where A is the amplitude of the step waveform and Fo is the clock frequency of the pseudo-random sequence generator.

For this application a clock frequency of 1MHz has been used, and from the equation above it can be seen that the frequency spectrum varies from that of a white noise source at frequencies approaching 1MHz, the deviation being 1½% at 100KHz. However this variation is taken into account in the noise measurement calculation and no error is introduced.

7. PRE-AMPLIFIER CONTROLLER

The Pre-amplifier Controller provides power supplies and control signals to the pre-amplifier, as well as generating other logic signals which control the operation of the noise equipment.

Logic signals are transferred to modules in the noise measuring equipment using a 16-bit control bus, and normally they are under the control of the micro-computer. However, occasionally it may be useful to control the equipment manually, and for this purpose thumb-wheel switches are provided on the Pre-amplifier Controller which can set control codes on the bus.

8. POWER SUPPLY

The importance of a low noise, mains operated power supply has been mentioned earlier. High levels of noise in the supply voltages to the pre-amplifier could degrade the noise measurement accuracy and limit the drain current operating range. In order to achieve sufficiently low noise and stable D.C. voltage supplies, it has been necessary to design a unit which fulfills the requirements of the analogue and digital circuitry within the noise measuring equipment. Voltage levels of +/-60V, +/-15V and +5V are supplied.

An internally screened mains transformer is equipped with three secondary windings. Each secondary winding is taken to an A.C. full-wave rectifier and a choke-input filter providing high attenuation to 100Hz mains ripple. Stabiliser circuits provide constant output voltages up to maximum current limits and incorporate circuitry for protection against an over-load or short-circuit condition.

An important feature of the equipment is the isolation of the earth return path from the logic circuitry, operated from +5V, from that of the analogue circuits, operated from +/-60V and +/-15V. This wiring scheme minimises the injection of noise from the logic circuits into the sensitive noise measuring amplifiers. Each module operates with two separate earth return paths which are joined only at a single common earthing point in the power supply.

Logic signals are transferred to modules in the noise measuring equipment using a 16-bit control bus, and normally they are under the control of the micro-computer. However, occasionally it may be useful to control the equipment manually, and for this purpose thumb-wheel switches are provided on the Pre-amplifier Controller which can set control codes on the bus.

9. TEMPERATURE CONTROLLED MEASUREMENTS

Cooled measurements may be performed by mounting the device in a cryostat assembly which is connected to the pre-amplifier by a co-axial cable. The device is mounted on a platform which is in contact through a known thermal resistance with a cold finger immersed in liquid nitrogen. The temperature of the device is controlled between 79°K and 350°K to within 1°K by applying heat to the platform through an electrical resistance, while monitoring the device temperature with a thermo-couple.

The Temperature Controller automatically sets the device to the required temperature and a unit similar to the one used here has been reported previously(7).

10. CONCLUSIONS

Investigation of the sources of low frequency noise in Fets and selection of devices for use in Infra-red detection and X and Gamma ray spectrometry require accurate noise measurements to be made on devices whose operating conditions can be accurately controlled over a wide range.

A computer controlled noise measuring equipment has been developed which can automatically measure the noise of Fets, NPN and PNP bipolar transistors in the temperature range 79°K to 350°K to an accuracy not previously attainable. Circuit design features such as dynamic filtering in the pre-amplifiers allows the devices operating conditions to be varied rapidly between noise measurements without loss of measurement accuracy.

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APPENDIX A

Some Typical Results

The accuracy of each noise measurement is to some extent dependent on the device characteristics and the current range over which it is being operated. A complete specification of the noise equipment and its measurement accuracy is given elsewhere(8). However, the absolute error on a noise measurement is typically less than 2% at 100KHz, increasing at low frequencies to 4% at 10Hz. The comparative error between different devices measured under the same conditions is typically less than 1% at 100KHz and 2% at 10Hz.

The table below shows the result of noise measurements performed on some typical devices at room temperature. The drain voltage is 4V in each case. The integration time is 100 seconds, giving a statistical standard deviation of 3½% at 10Hz, reducing to 0.07% at 100 KHz.

		NOISE IN $nV(Hz)^{-\frac{1}{2}}$				
		Frequency (Hz)				
Device	Drain Current	100K	10K	1K	75K	10
VX11012*	30mA	0.443	0.469	0.736	0.928	1.39
U309	10mA	1.09	1.11	1.12	1.38	2.14
2N4416	4mA	2.18	2.99	4.61	9.37	12.3
BF800	0.3mA	6.43	7.18	10.4	23.3	50.7

*Gate and substrate joined.

TABLE 1

Noise measurements performed on some typical devices with 4V drain voltage and 100 secs integration time.

APPENDIX B

Noise in the Drain Current Supply Circuit

The drain current may be considered, for purposes of noise calculation, in the first instance as being generated by applying a variable voltage V across a fixed resistor R1. The full-scale value, Im, of drain current is Vo/R1, where Vo is the maximum value of V.

The major noise source in the circuit is the thermal noise of R1, which injects a noise current In1 into the drain of the device under test. In1 is given by:-

$$In1 = (4KT/R1)^{\frac{1}{2}} \quad A(Hz)^{-\frac{1}{2}} \quad \ldots \quad (1)$$

Where K is Boltzmann's constant in JK^{-1}
and T is the temperature in degrees Kelvin.
R1 is in Ohms.

Noise in the Fet under test may be generated from a number of sources, and at low frequencies is strongly dependent on device operating conditions. However, at all frequencies the Fet noise will never be less than that due to the thermal agitation of carriers in the channel, which may be expressed as a noise current in the drain given by:-

$$Ino = (4KT\ gm\gamma)^{\frac{1}{2}} \ A(Hz)^{-\frac{1}{2}} \ \ldots \quad (2)$$

$$\text{and } gm = C.(Id)^{\frac{1}{2}} \ AV^{-1} \ \ldots \ \ldots \quad (3)$$

where gm is one mutual conductance of one device,
Id is the drain current supplied,
γ and C are constants.

Therefore, from equations 1, 2 and 3, the ratio of noise current from the Fet under test to the excess noise current from Ri is:-

$$Ino/In1 = Id^{\frac{1}{4}} \ (R1\,\gamma\ C)^{\frac{1}{2}} \ \ldots \ (4)$$

$$= Id^{\frac{1}{4}} \ (Vo\gamma\, C/Im)^{\frac{1}{2}} \ \ldots \ (5)$$

Thus the accuracy of the noise measurement is dependent on Vo and the drain current operating range, that is, the ratio of Id to Im. In order to allow a sufficiently wide range of drain current in each pre-amplifier while maintaining the noise measurement accuracy, it has been necessary to use a value of Vo of +60V.

In practice, the drain current is supplied through three resistors R6, R7 and R8 shown in figure 7. R8 supplies only a small current equal to the cascode current, and is sufficiently large for its noise contribution to be negligible. The majority of the drain current is supplied through resistor R7 which makes the most significant contribution to the noise signal. When operating at a drain current of 1/20th full-scale, R7 may increase the total noise measured by 5%. To increase the measurement accuracy at the lower limit of the drain current dynamic range, a first order correction is made in the noise calculation which subtracts the noise of R7 by estimating the mutual conductance of the device from the 100KHz noise measurement. This correction assumes that the equivalent noise voltage, En, of the device measured at 100KHz is largely that due to the

thermal agitation of carriers in the channel, and is given by:

$$En = (4KT\gamma/gm)^{\frac{1}{2}} \quad V(Hz)^{-\frac{1}{2}}$$

Other sources of noise in the drain current supply circuit are as follows:-

(i) noise from the +60V supply line,

(ii) the emitter resistance noise and base spreading resistance noise of T4 and T5,

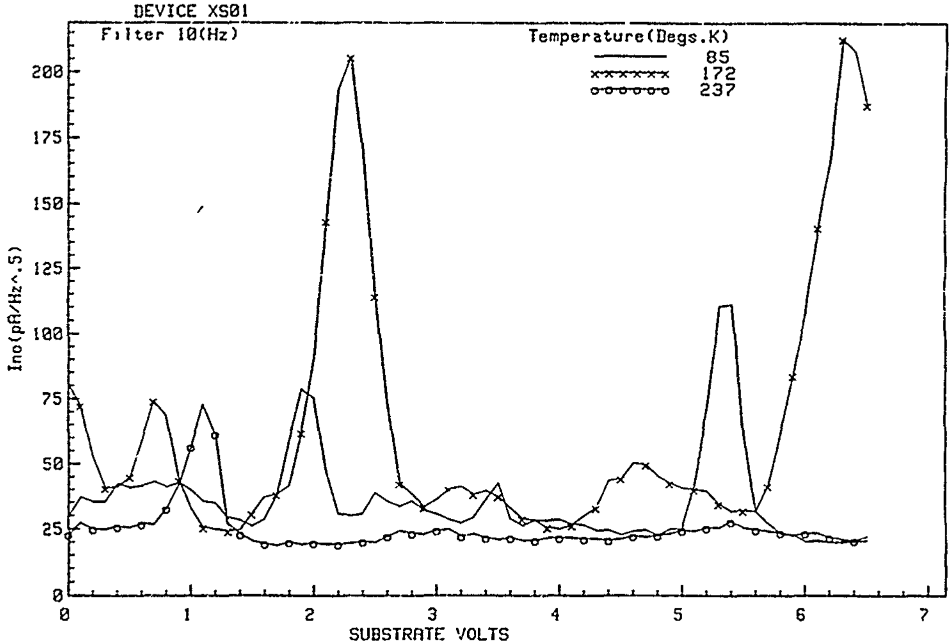
(iii) noise in T4 and T5 due to finite base current.

Noise from the +60V supply line may be injected into the drain through resistors R6, R7 and R8, and it is essential that the supply voltage is generated from a relatively low noise source. Additional filtering of the +60V line is provided in the pre-amplifier.

Capacitors C1, C2 and C3 provide a low impedance path for the noise from the filtered +60V supply line to the base of transistors T3 and T4; this serves to cancel the supply line noise in the currents through resistors R3, R6 and R7. The value of resistor R8 is sufficiently high for the noise current injected into the drain from the filtered +60V line to be negligible. R5 and C3 provide attenuation to any other noise voltage which may be present at the base of transistor T4 from whatever source.

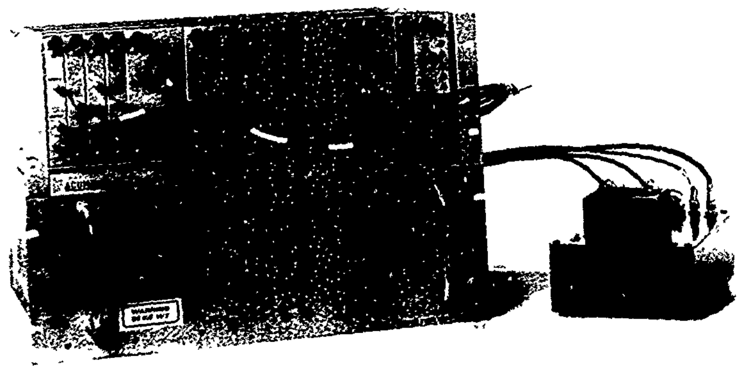
The base current noise from the transistor which supplies the drain current has been reduced by the use of a dual transistor configuration, T4 and T5, which requires only a very small base current in each transistor. The supply of additional current to the emitter of T4 through resistor R6 results in a lower emitter resistance noise in T4 than would be achieved with a true Darlington-pair circuit, although the base current noise in T4 is increased. The best compromise between these two noise sources in T4 is achieved by setting the value of R6 to be approximately 10 x R7. It has been found that the magnitude of the base current noise at a given collector current increases at low frequencies, typically an increase by a factor of five is observed between the partition noise measured at 100KHz and 10Hz. However, the use of a dual transistor configuration results in a negligible contribution from base current noise to the drain current supply under all operating conditions.

The contribution to the drain current supply from the emitter and base resistance noise of T4 and T5 is determined by the values of R6 and R7 respectively. The use of a supply voltage as high as +60V allows R6 and R7 to be sufficiently high for this noise contribution to be negligible.

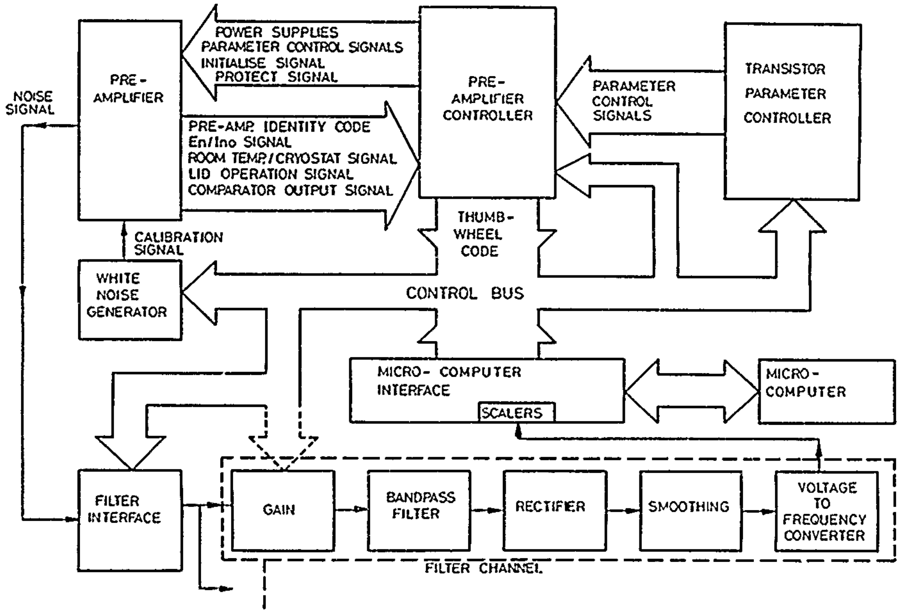


AERE - R 10642 Fig 1

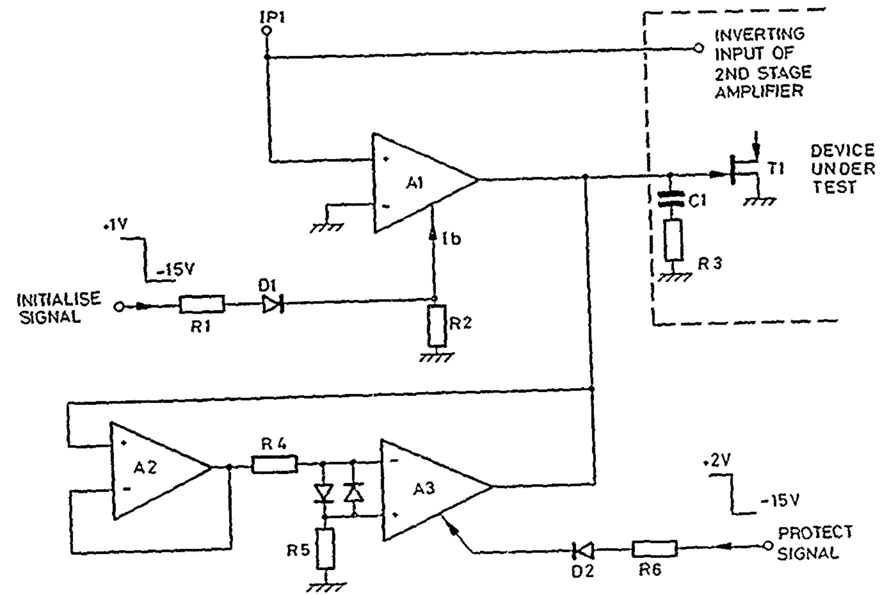
The variation of the low frequency noise of a JFet with substrate voltage and temperature



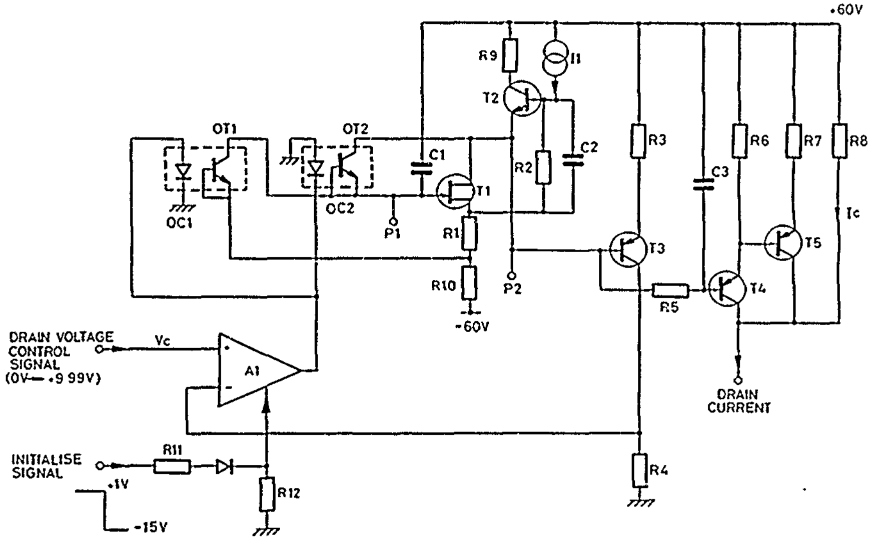
AERE - R 10642 Fig. 2
Photograph of the Noise Measuring Equipment



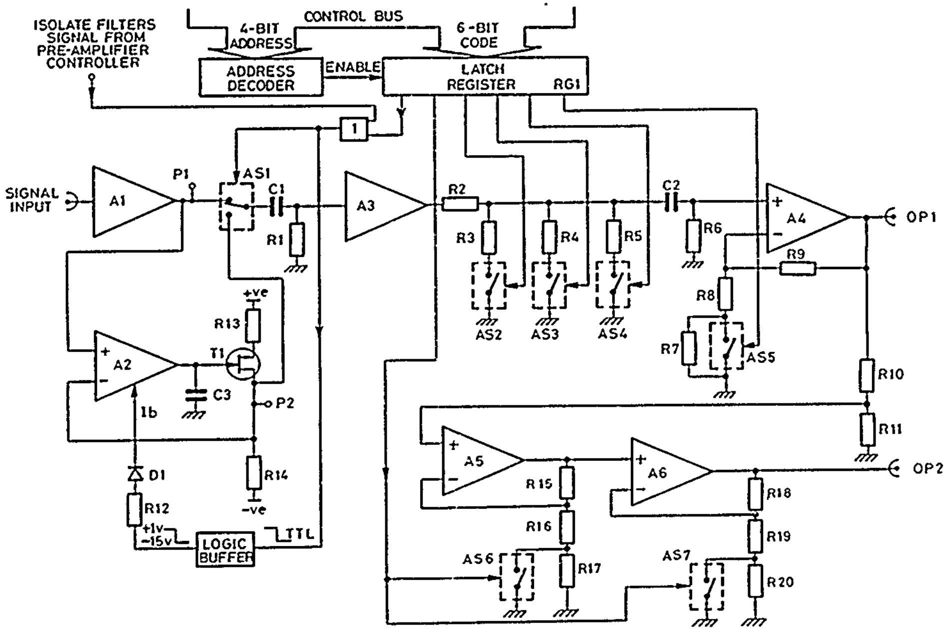
AERE - R 10642 Fig 3
Block diagram of the noise equipment



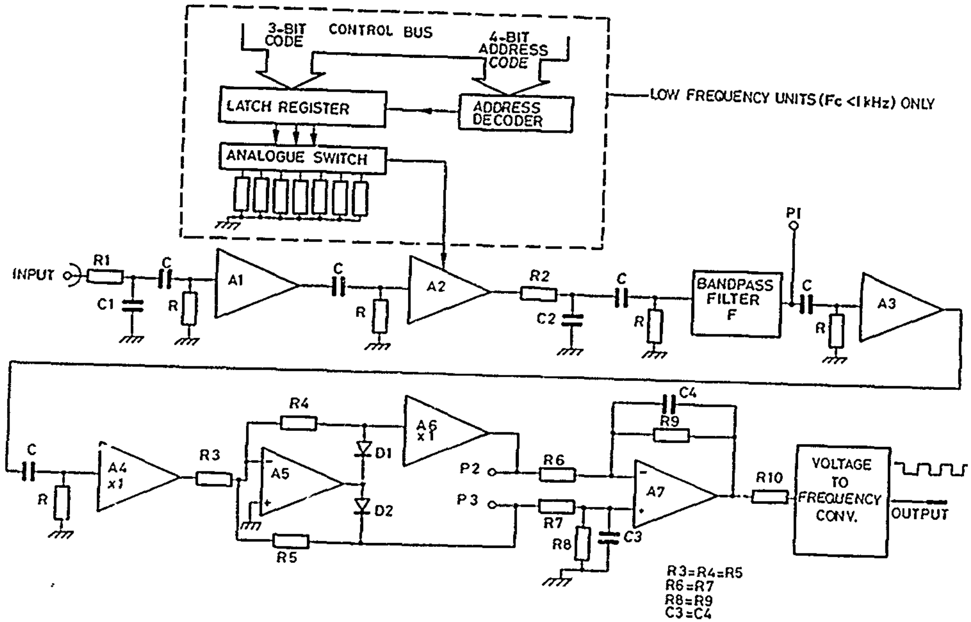
AERE - R 10642 Fig 6
Schematic diagram of the gate voltage control circuit



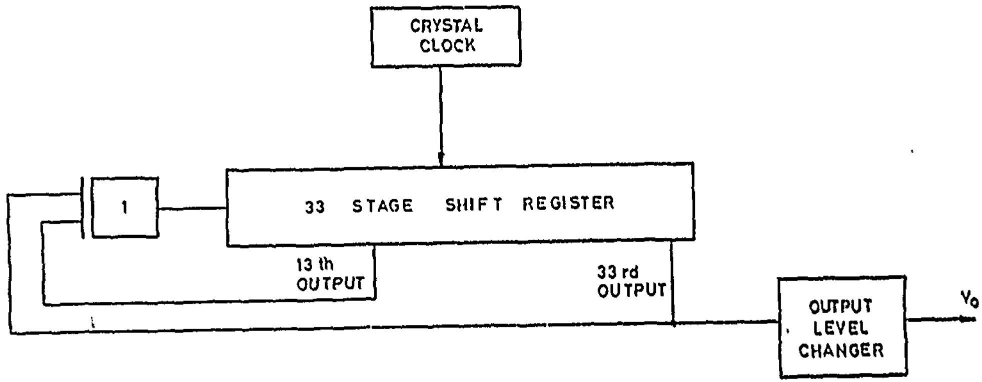
AERE - R 10642 Fig 7
Schematic diagram of the drain current control circuit



AERE - R 10642 Fig 8
Schematic diagram of the Filter Interface



AERE - R 10642 Fig 9
Schematic diagram of a filter channel



AERE - R 10042 Fig. 10
Schematic diagram of a pseudo-random sequence generator